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RESEARCH DEPARTMENT

REPORT

Field-store standards conversion:
amplifiers for ultrasonic quartz delay lines

No. 1970/24

Research Department, Engineering Division
THE BRITISH BROADCASTING CORPORATION

RESEARCH DEPARTMENT

FIELD-STORE STANDARDS CONVERSION: AMPLIFIERS FOR ULTRASONIC QUARTZ DELAY LINES

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(EL-41)

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Summary

A field store standards converter has been built which uses fused quartz delay-lines as the storage medium for the television signal. Amplifiers are required in order to recover the considerable losses in these delay-lines; the design of these amplifiers is described, with reference to the choice of the parameters of the system chosen for signal transmission and to the properties of the quartz delay-lines themselves.

1. Introduction

Many of the applications of ultrasonic delay lines in television broadcasting have been described,[1,2,3,4] and, in particular, the performance and specification of wide-band ultrasonic delay lines using fused natural quartz have been discussed.[5,6] In practice, the television signal is passed through the quartz as a modulated radio-frequency ultrasonic wave and appropriate terminal amplifiers are required in order to overcome the attenuation in the delay line and its electro-acoustic transducers. In most applications, the design of these amplifiers is straightforward but, where many delay lines are switched in cascade, as in a field-store standards converter,[7] the associated amplifiers must meet certain special requirements. The present report describes the delay line amplifiers used in a field-store standards converter; the design of the equalisers is described in a second report.[8]

2. Performance required

In general, the required amplifier performance depends on the modulation system adopted and this, in turn, depends on the performance of the delay line and on the requirements of the complete system. The field store standards converter requires a variable delay which is formed by switching up to 20 delay units in cascade and it is important that the unavoidable fluctuations in signal level which accompany this systematic switching process do not impair the converted picture.

A study has been made of the impairments[10] in a switched system using an amplitude-modulated signal and it was found that level changes of 0·06 dB and delay errors of 10 ns were just perceptible. The stringency of these demands on gain stability was avoided by choosing frequency modulation for the field store standards converter and relying on an efficient amplitude limiter to remove the cumulative effects of gain differences of up to 1 dB in each of the delay units.

The stability of the loss and time-delay of each quartz delay line is ensured by placing it in a stable temperature-controlled oven[11] and, although the usual tolerance for a quartz delay is ± 100 ns, timing errors are corrected in another part of the converter. Frequency-dependent irregularities in gain and delay will cause distortion of the f.m. signal and must be reduced to ± 0·2 dB and ± 4 ns by the equalisers;[8] most of these irregularities arise in the quartz but the amplifiers must clearly have the requisite stability of response.

The specified centre-frequency of the quartz delay lines used in the field store standards converter was 30 MHz, a compromise between the excessive dispersion at low frequencies and the excessive loss at high frequencies. The specified bandwidth was not less than ± 5 MHz, centred on 30 MHz and was limited partly by the performance of available transducers. It was hoped that a greater bandwidth would be realisable through equalisation but this was hindered by the spread of cut-off frequencies in the cascade of switched delay lines.

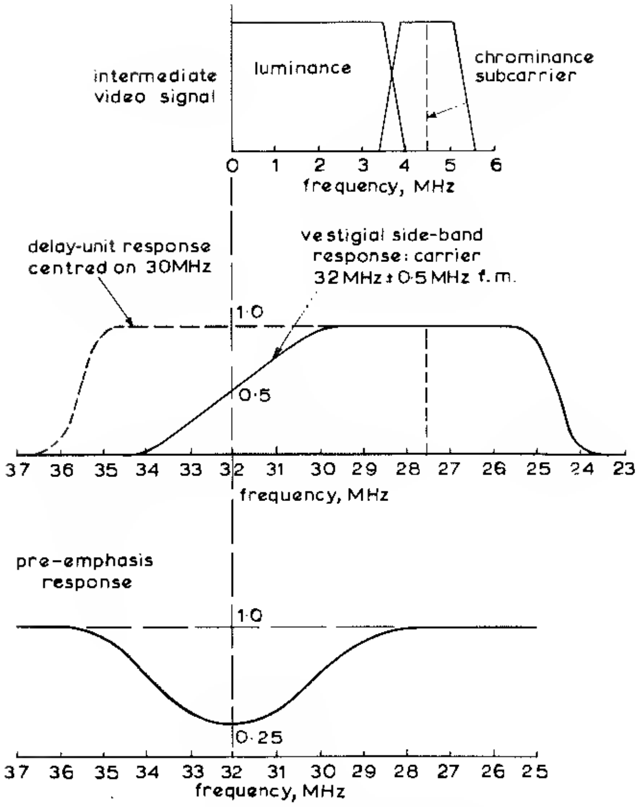


Fig. 1 • Transmission system responses

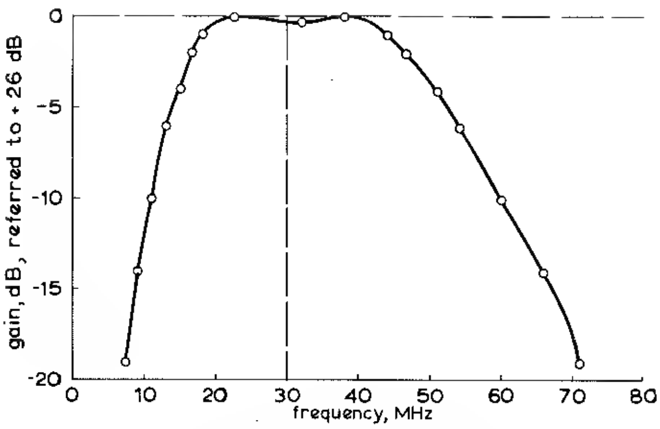


Fig. 2 • Input amplifier response

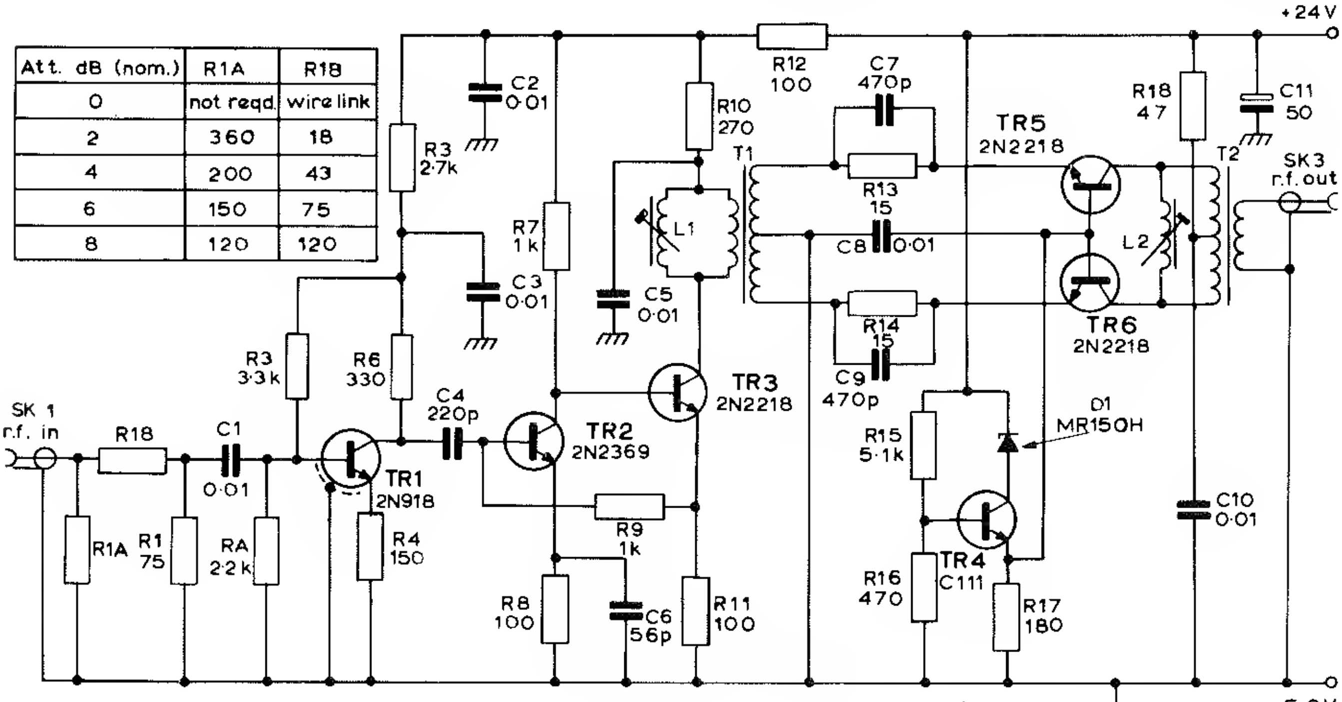
As the bandwidth available from the quartz delay-lines was not sufficient to accommodate a full double-sideband f.m. system, a low-deviation, vestigial-sideband system was adopted, using a quadrature-modulated (QAM) sub-carrier for the chrominance signal. Fig. 1 shows the essential features of the system.

The gain and signal-handling capacity of the amplifiers associated with each delay line will be described in the sections below and were determined by the following factors. First, a level of 0·5 V peak-to-peak was chosen as being convenient for the generation, detection and switching of the r.f. signal; second, the low r.f. signal level at the output transducer of the delay line was in all cases to be sufficiently in excess of the noise level in the first stage of the output amplifier to give an adequate signal-to-noise ratio. (See Appendix 1 and Section 5).

3. Amplifier preceding delay line

3.1. Input amplifier

The input amplifier raises the level of the 30 MHz input signal from 0·5 V peak-to-peak to 10 V peak-to-peak, with a response as shown in Fig. 2, for a load resistance of 75 ohms, which is convenient for test purposes; the optimum load is about 100 ohms.



Att. dB (nom.)	R1A	R1B
0	not reqd.	wire link
2	360	18
4	200	43
6	150	75
8	120	120

Fig. 3 • Input amplifier circuit diagram

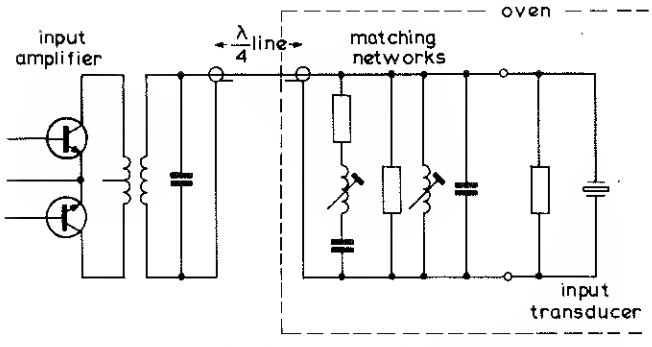


Fig. 4 • Matching of input amplifier to delay line

The gain may be reduced by a preset attenuator at the input terminal if the delay line has a low insertion loss. The circuit is fairly conventional and is shown in Fig. 3; TR4 supplies a low-impedance base-bias to the push-pull stage, TR5 and TR6, in order to stabilise the working point. The amplifier and its 24 V power supply occupy one, standard, 2-inch plug-in unit; the input and output connectors are isolated from the chassis to avoid multiple-earthing in the complete unit.

3.2. Input amplifier output stage

The amplified signal passes from the input amplifier into the high-temperature environment (up to 75°C) of the delay-line oven[11] and is applied to the input transducer of the delay line which may have a capacitance of up to 300 pF. It would seem desirable to include additional power-gain as close to the transducer as possible but this was found to be impracticable at this high temperature without restricting gain, bandwidth and linearity. Moreover, the additional heat dissipated inside the temperature-controlled environment of the delay line might jeopardise the stability of its delay-time. The output stage therefore consists of a passive network contained in a brass box (5 x 3 x 3 cm); the circuit is shown in Fig. 4. By damping and shunt-resonating the transducer capacitance at 30 MHz, a tuned circuit with a Q of 2 is formed and, with the addition of the series-resonant circuit, the admittance is transformed through the quarter-wavelength line to produce a load of approximately 100 ohms at the amplifier terminals. In practice, the quarter wave line is a coaxial cable of subminiature construction and designed to withstand the high temperatures in the oven.

The circuit has several useful features. The reactive compensation provided by the quarter-wave line and the series-tuned circuit enable the input amplifier to operate with a more nearly resistive load and also afford some degree of group-delay compensation for the transducer circuit. Adjustment of the components of the output stage circuit also gives some control over the amplitude/frequency performance at the transducer so that this may be fitted to the inverse of the response of each individual delay line.

A further point of interest concerns the connection to the transducer itself. Many early delay lines had transducer capacitance approaching the upper limit of 300 pF, but owing to the length of the internal connection between transducer and socket, the effective capacitance was much higher. It can be shown, however, that making the characteristic admittance of the connecting cable equal to the transducer susceptance ($Y_0 = \omega C$) results in the least increase in effective capacitance (see Appendix II). The solution adopted was to design the external network for a maximum capacitance of about 100 pF (a reactance of 50 ohms at 30 MHz), and to reduce the effective transducer capacitance to this value by fitting internal transformers at the transducers. Some manufacturers fit such transformers at the transducers and also arrange for the shunt reactance to resonate with the transducer capacitance at the mid-band frequency.

4. Amplifier following delay line

4.1. Output preamplifier

It was both desirable and practicable to include a low-noise pre-amplifier in the high-temperature environment, close to the output transducer of the delay line, in order to raise the level of the delayed (and attenuated) signal well above the level of noise and interference before it leaves the screening of the oven. A gain of 26 dB is required to raise the level from the 10 mV peak-to-peak output level from the delay line to 200 mV peak-to-peak, a suitable level for equalisation. The pre-amplifier circuit is shown in Fig. 5; a common-base stage is followed by a

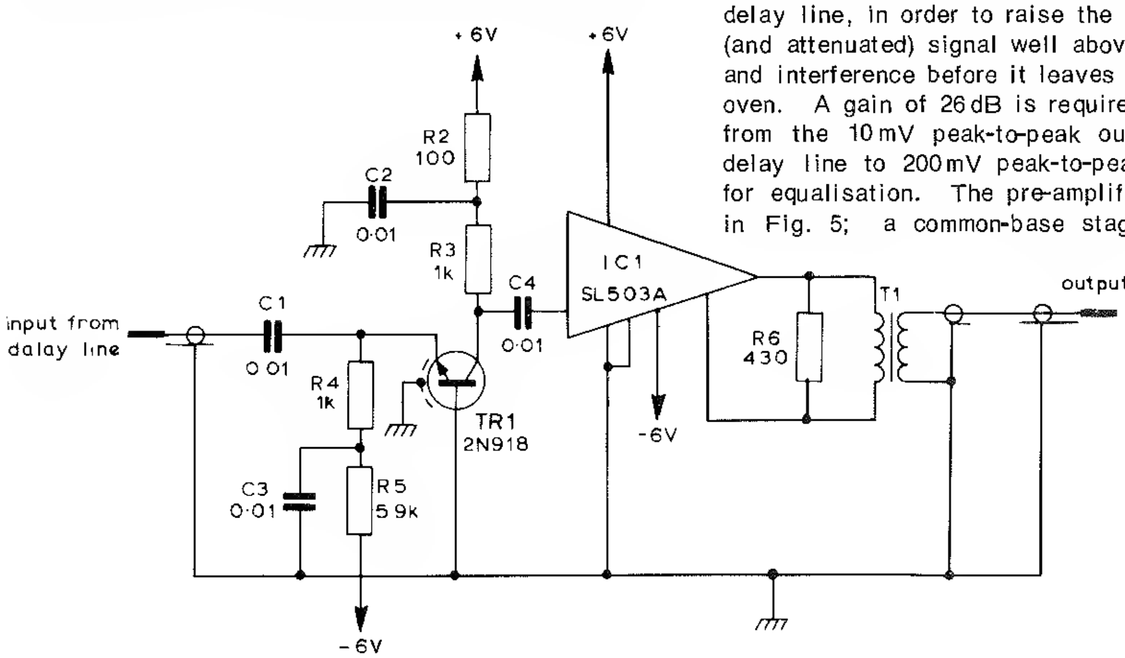


Fig. 5
Output pre-amplifier circuit diagram

linear integrated circuit amplifier housed in a brass box measuring 5 x 3 x 3 cm. The heat dissipation of about 120 mW introduces a slight non-uniformity into the temperature distribution of the delay-line environment but this is small and stable enough to be neglected.

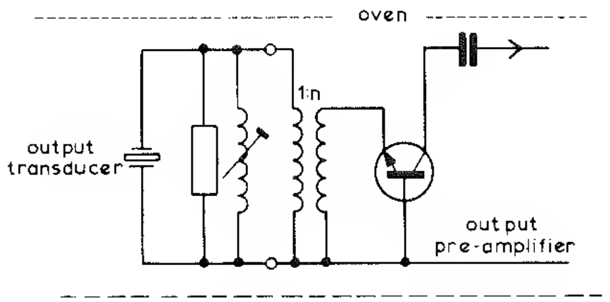


Fig. 6 - Matching of delay line to output pre-amplifier

Noise factor, bandwidth and gain are all dependent on the mis-match between the output transducer and the amplifier. Fig. 6 shows an equivalent circuit including a perfect transformer and Fig. 7 shows the effect of altering the transformer ratio. (The mid-band noise factor was measured, using a resistive source and the gain and bandwidth were calculated). As can be seen, there is a compromise to be made between high bandwidth and low noise factor. It was decided to give preference to bandwidth and to couple the transducer directly to the amplifier ($n^2 = 1$) since the majority of delay lines would have losses of less than 60 dB and the noise factor of 15 dB would be adequate (see Appendix 1).

In the equivalent circuit of Fig. 6, the transducer capacitance has been simply shunt-resonated with an inductor which may be introduced into the delay line assembly by the manufacturer or may be required in

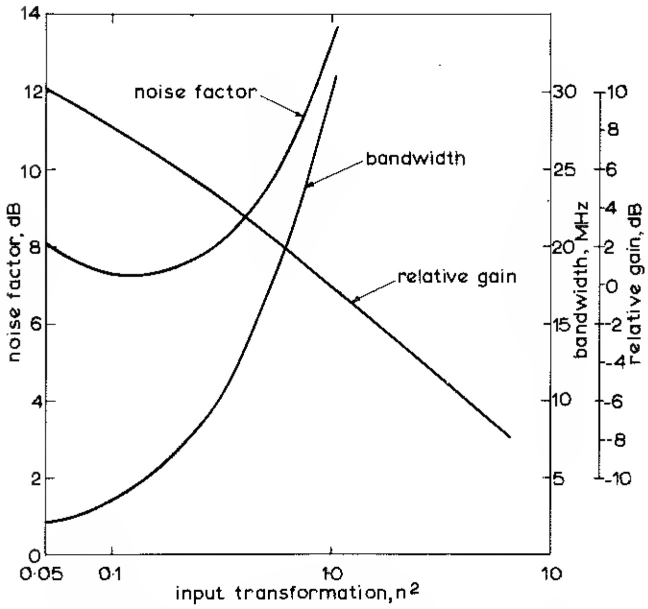


Fig. 7 - Gain, bandwidth and noise factor of output pre-amplifier

the pre-amplifier. (Some manufacturers prefer to introduce some series inductance in order to compensate for a poor high-frequency response. If this is the case, then the pre-amplifier must provide a 75 ohm load, either by series padding or by a transformer. In the interests of minimum group delay variation, however, it is preferable to omit the series components.)

4.2. Output amplifier, delay and gain equaliser

The signal emerges from the pre-amplifier in the oven at a maximum level of 0·2 V peak-to-peak, requiring a further amplification of 8 dB to reach 0·5 peak-to-peak so that the complete delay unit has a gain of unity. In order to cope with delay lines having insertion losses of extreme values, two alternative types of output amplifier have been designed, giving 8 dB and 18 dB gain, respectively, from 20 to 40 MHz, together with a preset attenuator section. Since a signal level of 0·2 V peak-to-peak is a convenient level at which to perform the necessary delay and gain equalisation for the unit, the output amplifier may contain, at its input, up to four equaliser sections, each of nominally unity gain; the alternative circuit arrangements are shown in Fig. 8.

The output amplifier is powered by a 12 volt supply, centre-tapped to earth by a zener diode circuit so as to provide 6-0-6 volts for the integrated circuit in the pre-amplifier. As with the input amplifier, the circuit earth is isolated from the chassis.

The output amplifier is first assembled without equalisers and its gain is adjusted to suit the loss in the delay line with which it is to operate.[12] At this stage, a detailed study is made of the required equalisation and the selection and adjustment of the equalisers is made empirically; this process is dealt with in detail in a separate report.[8]

5. Performance obtained

The signal level applied to the input of the delay line was well below the maximum voltage permitted for the transducer, but it would have been difficult to achieve the desired bandwidth at a higher voltage level. Nevertheless the signal-to-noise ratio of the delay units was adequate for the stringent requirements of the field store standards converter, even when using quartz delay lines of up to 60 dB insertion loss. Appendix 1 shows the results of calculation of the components of the output signal-to-noise ratio for a delay-unit having a 60 dB quartz loss. The r.f. pre-emphasis characteristic (shown in Fig. 1) was measured and found to modify the effective noise power spectrum from a parabolic to a roughly triangular distribution, for which the standard CCIR noise weighting law[13] gave a weighting of about 11 dB, leading to a calculated weighted output signal-to-noise ratio of 50 dB for one such delay unit. Assuming that the average number of cascaded delay units in the field-

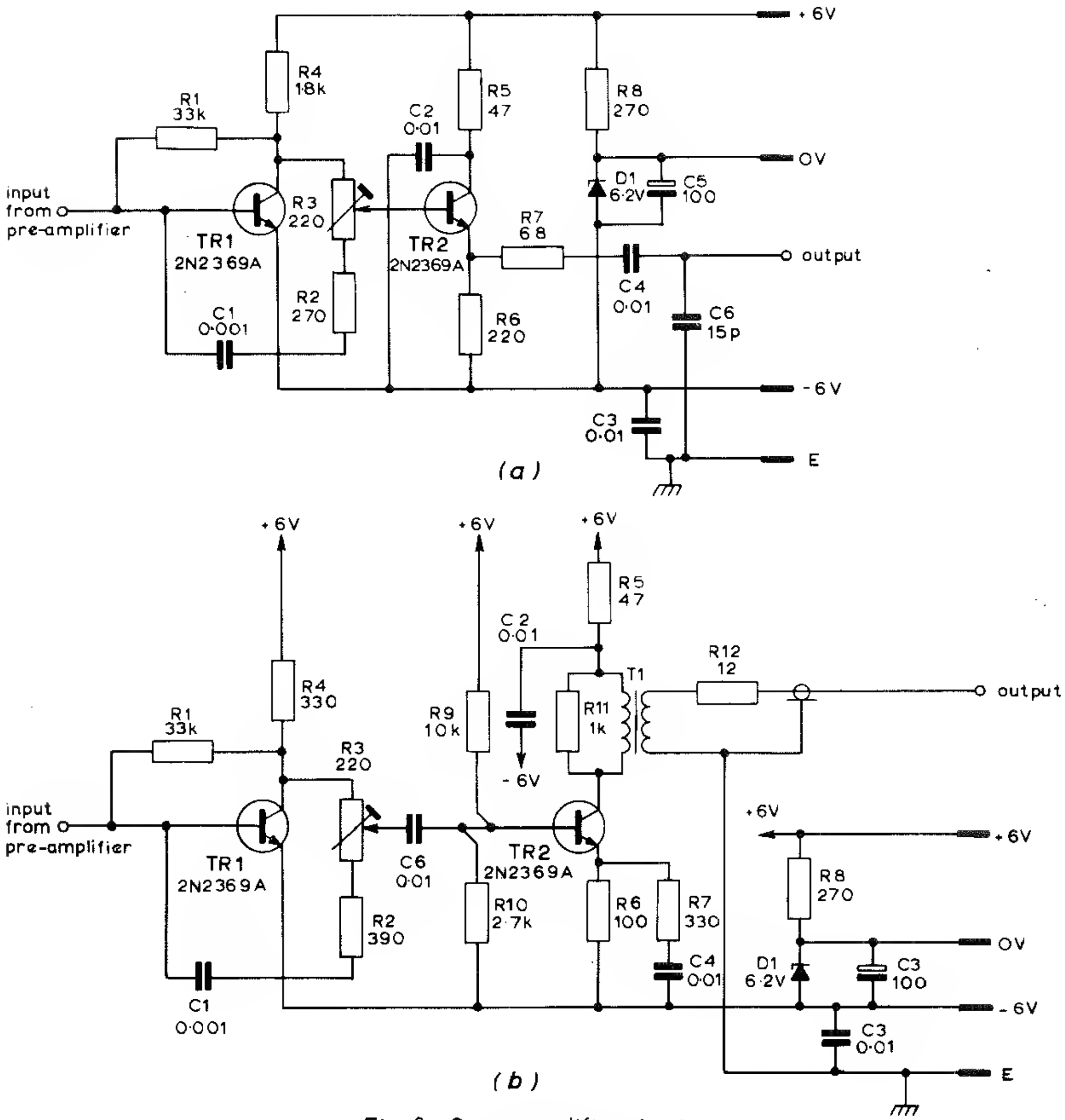


Fig. 8 - Output amplifier circuit

(*a*) 8 dB gain (*b*) 18 dB gain

store converter is twelve and the average quartz loss is 56dB, the expected value of the output signal-to-noise ratio for the whole converter would then be about 32dB r.m.s. or 43dB weighted. A value of about 44dB was assessed subjectively but, in this assessment, it was estimated that the major impairment was caused by secondary delayed signals.

Appendix 1 is, however, pessimistic for the following reason. It was assumed in the calculation that, owing to the low modulation index of the f.m. system, the r.f. pre- and de-emphasis were equivalent to video pre- and de-emphasis in respect of the noise spectrum. This is clearly not true for low modulation frequencies where the r.f. pre-emphasis will enhance the higher-order sidebands and therefore improve the signal-to-noise ratio.

Since the amplifiers were working near to the limit of their power-handling capability, their linearity left something to be desired; while this was adequate for the chosen f.m. system using a QAM chrominance subcarrier, it would not have been adequate for a system transmitting the colour signal by frequency-division multiplex which would give rise to interfering intermodulation products.

The system bandwidth was limited by the quartz delay lines having a bandwidth, at −3dB, of 10MHz; the amplifiers themselves had sufficient bandwidth to allow all the delay units to be equalised to within ± 0·2dB over this range. Secondary signal[6] levels of less than −46 dB had been specified for the delay lines themselves; it was important, therefore, that direct breakthrough (i.e. from input to output) was minimised.

By careful attention to the construction of the amplifiers the direct breakthrough was found to be satisfactory, indicating an isolation of over 106 dB between the input amplifier and the output pre-amplifier.

6. Conclusions

Amplifiers have been designed and constructed to supply a 0·5 V peak-to-peak signal at 30 MHz ± 5 MHz to an ultrasonic delay line and to recover the delayed signal, forming a unity-gain, equalised delay unit. The performance of the amplifiers was adequate for the requirements of the field store standards converter.[7] In particular, the linearity was adequate for the single-sideband f.m. system adopted in the converter, using a quadrature-amplitude-modulated chrominance subcarrier. Should future delay lines become available with much greater bandwidth, the adoption of a frequency-division multiplex analogue system would demand greater linearity.

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APPENDIX I

Signal-to-Noise Ratio

The factors contributing to the signal-to-noise ratio are listed below in the form of a balance-sheet. The last item on the list is the weighted, output signal-to-noise ratio, for which a value of 50 dB is required in order to balance the sheet. The preceding figures are all derived from system design parameters, delay-line specification limits, noise-factor measurements and calculations of noise power spectra and are described in the report and discussed in Section 5. Noise in f.m. systems is described in Reference 14.

Gains			
Input signal level		+131 dBμV	
FM/AM Improvement			
F.M. Index	−20 dB		
Phase mod noise	+ 5 dB		
De-emphasis	+ 9 dB		
CCIR weighting	+11 dB		
VSB	− 2 dB		
Signal-to-noise improvement		+3 dB	
Equivalent input signal level for AM system			+134 dBμV
Losses			
Thermal noise level	+ 9 dBμV		
Noise factor of pre-amp	+15 dB		
Signal loss in delay line	+60 dB		
Equivalent input noise level		+84 dBμV	
Required output signal-to-noise ratio		+50 dB	
Equivalent input signal level for AM system			+134 dBμV

APPENDIX II

Effective Capacitance of Untuned Transducer

A transducer capacitance of susceptance B is transformed through a transmission line of length θ and characteristic admittance Y_0 to give a susceptance

$$B' = Y_0 \frac{B + Y_0 \tan\theta}{Y_0 - B\tan\theta}$$

so that the increase in susceptance is

$$B' - B = \frac{(Y_0^{\,2} + B^2)\tan\theta}{Y_0 - B\tan\theta}$$

If θ is small, this increase may be minimised by making $(Y_0^{\,2} + B^2)/Y_0$ minimal, i.e. by making $Y_0 = B$. Then

$$B' - B = Y_0 \frac{2\tan\theta}{1 - \tan\theta}$$

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ADDENDUM

RESEARCH DEPARTMENT – BRITISH BROADCASTING CORPORATION

Research Department Report No. 1970/24

FIELD STORE STANDARDS CONVERSION:
WIDEBAND MATCHING FOR ULTRASONIC QUARTZ DELAY LINES

Summary

This addendum shows how the overall bandwidth of the delay-lines in the 625/50 to 525/60 standards converter (C06/508) has been increased from 10 MHz to 14 MHz. This increase has enabled the modulation system within the converter to be changed from vestigial-sideband f.m. to double-sideband f.m., and the deviation to be doubled, thus improving the overall signal/noise ratio of the converter.

1. Introduction

The 625/50 to 525/60 field-store standards converter uses about thirty cascaded delay units, most of which are switched systematically in and out of circuit during the line and field-blanking intervals. As described in the main report (1970/24), each delay unit comprises a fused quartz delay-line together with amplifiers and equalisers which are adjusted to provide unity gain and a uniform amplitude response in the pass-band. The propagation properties of the quartz are such that it operates best at about 30 MHz and the video signals pass through the cascaded delay-lines in the form of a frequency-modulated, 30 MHz carrier. When the 525/60-to-625/50 field-store converter (C06/506) was first developed, it was considered that the bandwidth available from the quartz delay-lines was not sufficient to accommodate a double-sideband (d.s.b.) f.m. system and consequently a vestigial sideband (v.s.b.) system was adopted. Later lines, however, were found to have a wider bandwidth but it could not be utilised because of the design of the matching network at the delay-line input. Redesign of this network has permitted the overall bandwidth to be made large enough to accommodate a double-sideband f.m. system with greater deviation, thus improving the signal/noise ratio (notably of the chrominance signal) and simplifying the design of certain other units such as the chrominance a.g.c. unit. This addendum describes the new matching network and outlines the revised wider-band modulation system.

2. The system bandwidth

The v.s.b. system originally adopted is described in Section 2 of the main report. The essential features of that system, together with those corresponding to the new d.s.b. system, are shown in Fig. 9. It is evident that, to permit the new system, the delay lines need to operate over the band 24 – 38 MHz, compared to 25 – 35 MHz for the original system.

3. The wideband matching network

As described in Section 3.2 of the main report (1970/24), the signal from the input amplifier passes through a fairly long cable (about 1 metre) into the delay-line oven and is applied to the input transducer of the delay line via a matching network. The restricted bandwidth of the original arrangement was due mainly to variation of impedance of the amplifier load over the frequency-band. The new matching network presents a substantially-constant 75Ω resistive load impedance to the amplifier over the band 24 – 38 MHz.

The equivalent circuit of the input transducer is given in Fig. 10. The component values vary widely from one delay line to another, but C generally lies in the range 50 to 250 pf, and R in the range 400Ω to 4 kΩ. The inductance L is usually present (and, if so, it resonates C at some frequency within the range 25 – 35 MHz) but is absent on some lines.

The transducer is resonated at the mid-band frequency of 30 MHz* by adding either shunt inductance or shunt capacitance as required. Expressed as an impedance, the equivalent resistance and reactance of this shunt combination vary over the frequency band as sketched in Fig. 11; the negative reactance slope may be corrected by adding a series-tuned

* The fractional bandwidth is fairly large and therefore the geometric mid-band frequency is used rather than the arithmetic mean of the band limits.

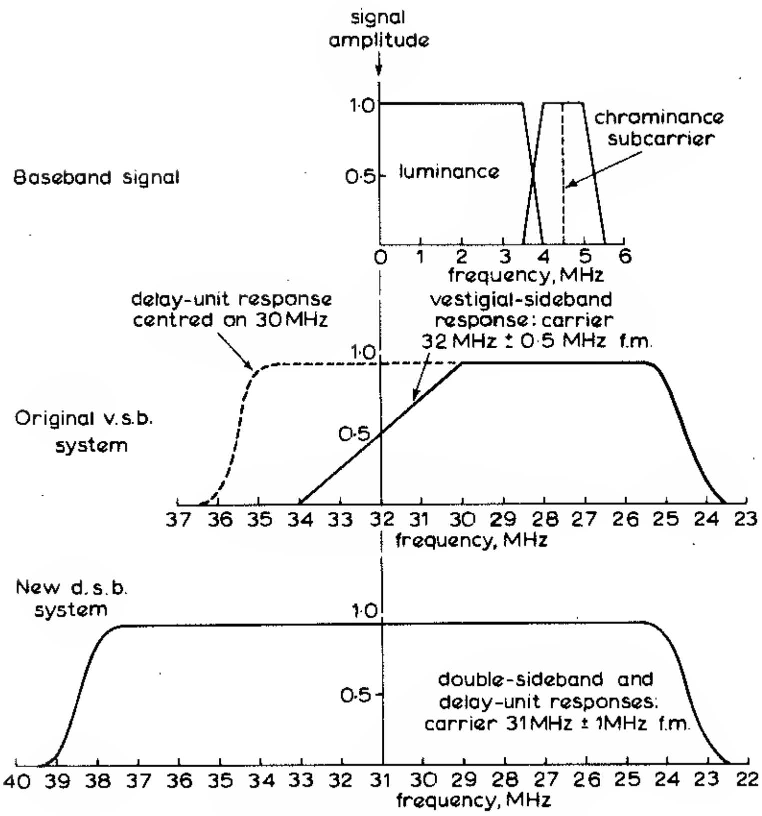

Fig. 9 - Transmission system responses

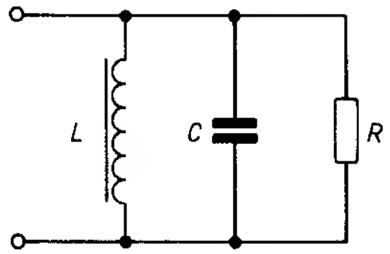

Fig. 10 - Equivalent circuit of the input transducer

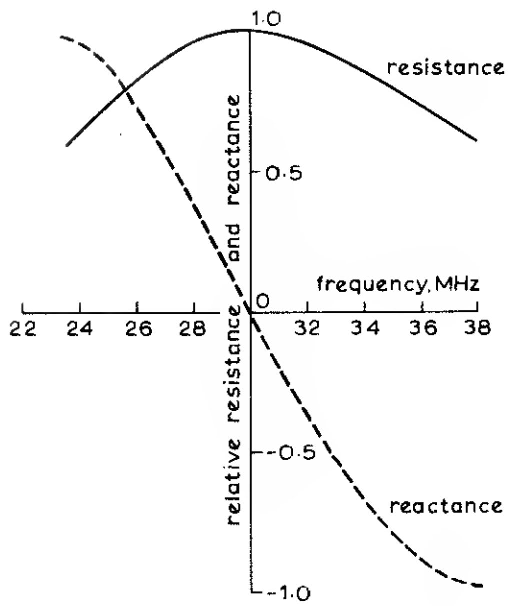

Fig. 11 - Impedance characteristics of the input transducer

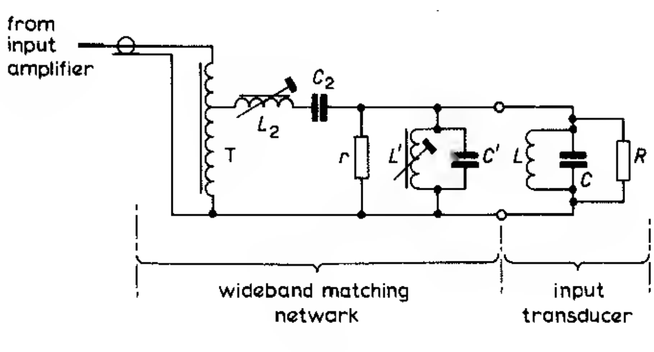

Fig. 12 - The wideband matching network

circuit (tuned to 30 MHz) in series, but the variation of resistance remains. It is shown in the Appendix that this variation depends upon the product CR and that, for the mismatch loss at the output of the input amplifier to be less than 0·1 dB over the band, CR must not exceed 7·55 ns.

Thus, depending upon the value of C, a damping resistor of suitable value must be connected across the input transducer. For most delay lines, the impedance level is then less than 75 ohms; thus a wideband transformer is required to match the 75-ohm cable connecting the network to the input amplifier.

The circuit of the wideband matching network takes the form shown in Fig. 12. R, C and L represent the constants of the input transducer, r is the damping resistor and either L' or C' is added to resonate the whole at 30 MHz; L_2 and C_2 represent the series tuned circuit which corrects the resulting negative reactance slope and T is the wideband transformer. In a practical case, having measured the resonant frequency and the values of C and R for a given transducer, all the component values of Fig. 12 are deduced from Table 1 in the Appendix; the resulting reflection coefficient is then less than 10%, and hence the mismatch is less than 0·1 dB, over the band 24 to 38 MHz.

4. Signal-to-noise ratio

In the original arrangement, the input transducers were damped with 75 ohms. Thus the new technique would not affect the r.f. signal-to-noise ratio were it not for the fact that the network usually needs to include a transformer; because of this, the signal level (and hence the signal-to-noise ratio) is reduced. It has been found possible, however, to restore the signal to its original level, on all but those few delay lines whose input capacitance and insertion loss are both very high, by reducing the attenuation previously inserted at the input to the samplifier. Thus the wideband matching technique has not affected the r.f. signal-to-noise ratio, except for the above few cases – where it was worsened by up to 3 dB. The associated changes from v.s.b. to d.s.b. working with increased deviation has, of course, improved the overall video signal-to-noise ratio.

5. Conclusions

By providing a matching network of the type shown in Fig. 12 at the input transducer of each delay line it has been possible to increase the overall bandwidth to 14 MHz. This has enabled the modulation system to be changed from vestigial-sideband to double-sideband f.m., and the deviation to be doubled; as a result the overall signal-to-noise ratio of the standards converter has been improved.

6. Appendix to Addendum – Derivation of component values in Fig. 12.

Say that the calues of R and C (see Fig. 10) and the resonant frequency of the transducer are known. Then either L' or C' is added to resonate the transducer at 30 MHz, and the damping resistor r is added (see Fig. 12). Say these components are, together, equivalent to L_1, C_1 and R_1 in parallel. The admittance of this combination is given by

$$Y = 1/R_1 + j\omega_0 C_1 (\omega/\omega_0 - \omega_0/\omega) \qquad \text{where } \omega_0{}^2 = 1/L_1 C_1$$

$$\approx 1/R_1 + j2\delta C_1 \qquad \text{where } \delta = (\omega - \omega_0) \text{ and } \delta \ll \omega_0{}^*$$

and its impedance by

$$Z = R' + jX' = 1/Y = R_1/(1 + 4\delta^2 C_1{}^2 R_1{}^2) - j2\delta C_1 R_1{}^2/(1 + 4\delta^2 C_1{}^2 R_1{}^2)$$

Say x and R_0 are chosen such that

$$R' = xR_0 \text{ at } \delta = 0$$

and $R' = R_0/x$ at $\delta = \pm\,\Delta$

$$\text{Hence } R_0 = (x^2 - 1)^{1/2}/2\Delta C_1 x \tag{1}$$

and, if the characteristic impedance of the cable connecting the input amplifier to the matching network is Z_0, the turns ratio (n) of the transformer in the matching network (see Fig. 12) is given by

$$n = (Z_0/R_0)^{1/2} = (2Z_0 \Delta C_1 x)^{1/2}/(x^2 - 1)^{1/4} \tag{2}$$

* Strictly, δ is not very small compared to ω_0 over the whole bandwidth but this approximation is sufficiently accurate for the present purpose.

At $\delta = \pm\Delta$, the series compensation required is $\pm(x^2 - 1)/2x^2\Delta C_1$.

Now the reactance of the series combination L_2C_2 (see Fig. 12) is given by

$$X_2 = \omega_0 L_2(\omega/\omega_0 - \omega_0/\omega) \qquad \text{where } \omega_0^{\,2} = 1/L_2C_2$$

$$\approx 2\delta L_2 \qquad \text{where } \delta = (\omega - \omega_0) \text{ and } \delta \ll \omega_0$$

$$\therefore \; L_2 = (x^2 - 1)/4x^2\Delta^2 C_1 \qquad (3)$$

In the above equations x is the standing-wave ratio (SWR) on the cable connecting the input amplifier to the matching network, both at the band centre ($\delta = 0$) and at the limits of the band ($\delta = \pm\Delta$). At intermediate frequencies the SWR is less than x. It is reasonable for x to have any value within the range 1 to 1·2 since the reflection coefficient at the output terminals of the input amplifier will then be less than 10%, and the mismatch loss less than 0·1 dB, over the total bandwidth 2Δ. Ideally n should be allowed to have any value, greater or less than unity, so as to achieve the best possible signal/noise ratio for a given value of C_1. Thus, for this application, n should be less than unity if C_1 is less than 84 pf and greater than unity if C_1 is greater than 84 pf. There were, however, so few delay lines for which the value of C_1 was less than 84 pf that it was unnecessary to consider values of n less than unity. Values of n had also to be restricted to the simple ratios typical of wideband transformers; the value of x was accordingly adjusted, having chosen the best transformer ratio, such that the reflection coefficient did not exceed 10%. The resulting values of R_1, C_2, n and the maximum reflection coefficient over the band 24 to 38 MHz (calculated from Equations (1), (2) and (3) with the above constraints) are given as a function of C_1 in Table 1.

TABLE 1

Component Values of Input Matching Network as a Function of Input Transducer Capacitance

C_1 (pf)	R_1 (ohms)	C_2 (pf)	n	Maximum reflection coefficient (%)
40	78	125	1	2·0
50	79·8	100	1	3·0
60	81·5	83·5	1	4·3
70	84·3	71·8	1	6·2
80	88·0	62·5	1	8·7
90	52·0	134	5/4	4·0
100	53·0	121	5/4	5·1
110	54·0	110	5/4	6·1
120	55·9	101	5/4	8·0
130	48·0	122	4/3	7·0
140	49·2	114	4/3	8·4
150	37·5	170	3/2	6·1
160	37·6	160	3/2	7·0
170	38·3	150	3/2	7·6
180	39·0	142	3/2	8·7
190	30·0	204	5/3	5·8
200	30·5	194	5/3	6·7
210	31·0	185	5/3	7·6
220	31·6	176	5/3	8·5
230	32·1	169	5/3	9·5
240	32·6	162	5/3	10·3

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